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DESIGN OF BANDPASS FILTERS WITH PARALLEL THREE-LINE COUPLED MICROSTRIPS

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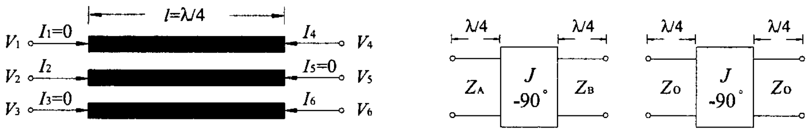
Bandpass filters are designed based on parallel coupled three-line microstrip structures. The relation between the circuit parameters of a three-line coupled microstrip section and an admittance inverter is established. A design graph for substrate with $\varepsilon_r = 2.2$ is provided. Two Chebyshev filters, of order 2 and 1dB passband ripple, with fractional bandwidths 40% and 60% are fabricated and measured. Good agreement between prediction and measurement is obtained.

1 Introduction

Parallel coupled-line microstrip filters has been found to be one of the most commonly used microwave filters in many practical wireless systems for several decades [1]-[2]. In addition to the planar structure itself and relatively wide bandwidth, the major advantage of this kind of filters is that its design procedure is relatively simple. Based on the insertion loss method [1], filter functions of maximally flat and Chebyshev type can be easily synthesized. In addition, the filter performance can be improved in a straightforward manner by increasing the order of the filter.

Bandpass filters with relatively wide bandwidths are useful in many applications. When these filters are to be realized by parallel coupled microstrip lines, one of the main limitations can be the small gap sizes of the first and the last coupling stages. The more fractional bandwidth, the smaller gap size is required to increase the coupling. Obviously, shrinking the gap size is not the only way to increase coupling of coupled lines. It is the idea of our work [3] to construct a coupled section with a symmetric three-line microstrip structure shown as in Fig.1. It can be shown quantitatively that a coupled three-line section has larger coupling than a traditional two-line coupled section with the same gap size. This implies that the tight gap size for designing a wideband filter can be released if a three-line structure is used instead of a two-line section.

Fig.1 Cross-section view of a parallel coupled three-line microstrip.



(a) (b)

Fig.2 Reduction of a coupled three-line section to a two-port network. (a) A coupled three-line section as a six-port network. (b) Equivalent admittance inverter. (c) Further approximated admittance inverter.

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2 Equivalent circuit of a three-line coupled section

There are three quasi-TEM modes in the parallel coupled three-line microstrip structure. The circuit parameters of a coupled section constructed of three-line microstrip can be obtained as follows. Let $[L]$ and $[C]$, respectively, be the inductance and capacitance matrices per unit length for the structure. The two matrices can be calculated by invoking the spectral domain approach (SDA) [4]. It is known that the eigenvectors of the matrix product $[L][C]$ form the eigenvoltage matrix $[M_V]$ for the three dominant modes. Due to the symmetry of the structure, $[M_V]$ can be written as

$$[M_V]=\begin{bmatrix} 1 & 1 & 1 \\ m_1 & 0 & -m_3 \\ 1 & -1 & 1 \end{bmatrix} \tag{1}$$

A section of coupled three-line microstrip is a six-port network as shown in Fig.2(a). The relation between the port voltages and port currents can be established via the impedance matrices $[Z_a]$ and $[Z_b]$:

$$\begin{bmatrix} V_a \\ V_b \end{bmatrix}=\begin{bmatrix} Z_a & Z_b \\ Z_b & Z_a \end{bmatrix}\begin{bmatrix} I_a \\ I_b \end{bmatrix} \tag{2}$$

where $[V_a] = [V_1, V_2, V_3]^T$, $[V_b] = [V_4, V_5, V_6]^T$, $[I_a] = [I_1, I_2, I_3]^T$, and $[I_b] = [I_4, I_5, I_6]^T$. The impedance matrices $[Z_a]$ and $[Z_b]$ are found to be

$$[Z_a] = [M_V]\ \mathrm{diag}[-jZ_{mi}\cot\theta_i]\ [M_V]^{\mathrm{T}} \tag{3}$$

$$[Z_b] = [M_V]\ \mathrm{diag}[-jZ_{mi}\csc\theta_i]\ [M_V]^{\mathrm{T}} \tag{4}$$

where $\theta_i = \beta_i l$ with β_i being the phase constant of the i^{th} mode and l the length of the coupled section. In (3) and (4), Z_{mi} is given as

$$Z_{mi}=\frac{Z_{oi}}{m_i^2+2} \tag{5}$$

where Z_{oi} is the characteristic impedance of mode i, and $m_2 = 0$.

Making the coupled three-line section in Fig.2(a) equivalent to a two-port, let port 2 be the input port and ports 4 and 6 be the output port. Following the approximations used in [3,5] to establish the equivalence between the coupled line section in Fig.2(a) and the inverters in Fig.2(b) and Fig.2(c), we can obtain

$$m_1Z_{m1} - m_3Z_{m3} = J\,Z_o^{\,2} \tag{6a}$$

$$m_1Z_{m1} + m_3Z_{m3} \approx Z_o\,(J^2Z_o^{\,2} + 1) \tag{6b}$$

This is the key step in our design procedure which greatly simplifies the determination of line width and line spacing of each coupled section in a general N^{th} order three-line bandpass filter. According to the design equations for a bandpass filter, the value of JZ_o for each admittance inverter of a general N^{th} order filter can be simply determined from the values of lumped circuit elements of the low-pass filter prototype. Once JZ_o is known, (6a) and (6b) can be solved simultaneously to determine the values of m_1Z_{m1} and m_3Z_{m3} for each coupled three-line section. The line width and spacing for each coupled three-line section can then be determined with the aid of the design graph as shown in Fig.3, where the substrate $\varepsilon_r = 2.2$. The design graph should be replaced if a substrate with different value of ε_r is used.

3 The fabricated filters and measurements

Two filters are designed and fabricated as illustrative examples. The filters have order $N = 2$, 1 dB passband ripple, and center frequency $f_o = 5.8$ GHz. The fractional bandwidths are 40% and 60%. The pattern of the circuit layout is shown in Fig.4. It is to be noted that a power combining circuit is required at the load end, to properly combine the signals on the first and the third lines, and to deliver the combined power to the load. This circuit should also correct the impedance mismatch since the load impedance for an even-order Chebyshev low-pass prototype is not unity.

Before fabricating the circuit, we have to compensate the parasitic capacitances resulted from the open-ends of each resonator and the series gap coupling between neighboring conductors. It is found that the latter is negligible after the former has been eliminated by a foreshortening at both ends of each line section.

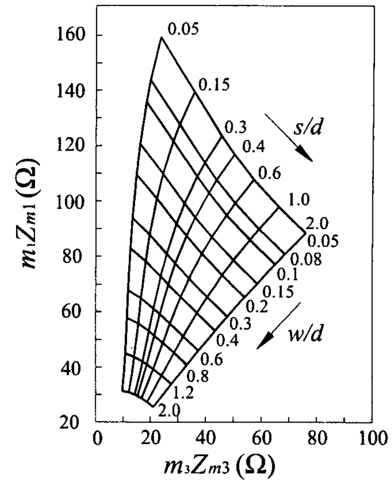

Fig.3 Design graph for a three-line microstrip, $\varepsilon_r = 2.2$.

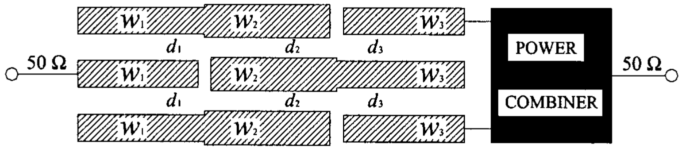

Fig.4 Circuit layout for a three-line Chebyshev filter of order 2.

We use the full-wave simulator IE3D [6] to validate our design before the circuit is fabricated. Figs.5 and 6 show the simulated and the measured S-parameters results for the filters with $\Delta = 40\%$ and 60%, respectively. A substrate with $\varepsilon_r = 2.2$ and thickness $d = 0.508$ mm is used. By coincidence, the three sections for each filter have approximately the same line width w/d and line spacing s/d. The length of each coupled section is 9.89 mm, which is about a quarter of guided wavelength at the center frequency. The measurement is performed using an HP8720 network analyzer. Both the $|S_{11}|$ and $|S_{21}|$ frequency responses have a good agreement between the measurement and simulation.

4 Conclusion

A systematic design procedure is given for synthesizing parallel coupled line filters based on a three-line microstrip system. The key step is to establish the equivalence between a coupled three-line microstrip section and an admittance inverter circuit. With each coupled section being compensated by the open-end discontinuities, a good agreement between the predicted and measured results is obtained.

Acknowledgements

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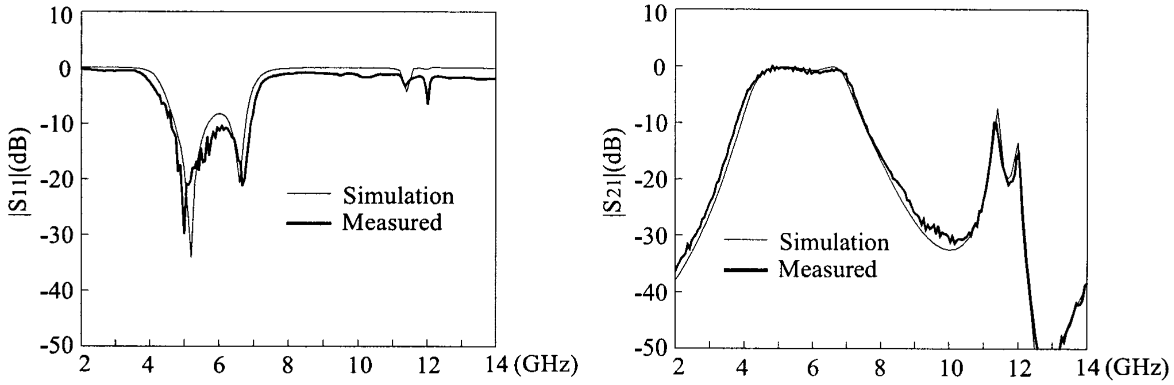
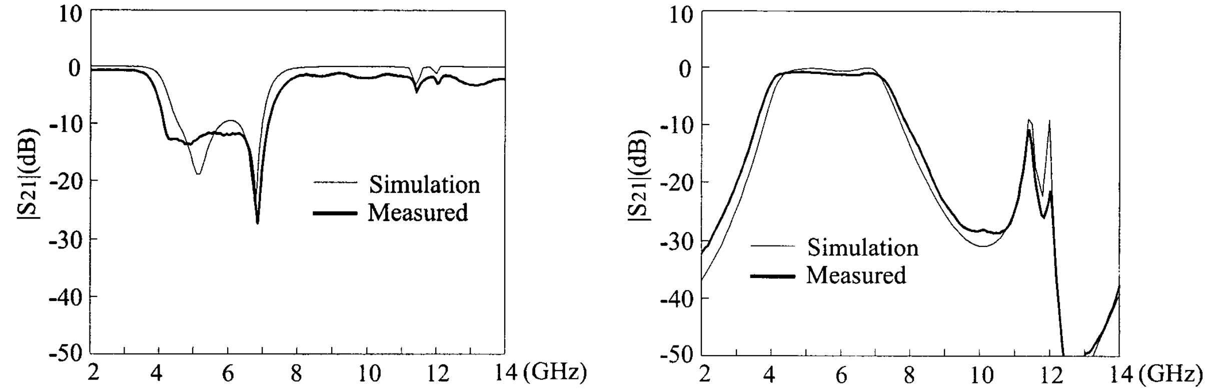


(a) (b)

Fig.5 Measured and simulated responses for a three-line filter with f_c = 5.8GHz and Δ = 40%. (a) $|S_{11}|$. (b) $|S_{21}|$. Parameters: d = 0.508mm, ε_r = 2.2, $w_1/d \approx w_2/d \approx w_3/d = 0.934$, $s_1/d \approx s_2/d \approx s_3/d = 0.321$.



(a) (b)

Fig.6 Measured and simulated responses for a three-line filter with f_c = 5.8GHz and Δ = 60%. (a) $|S_{11}|$. (b) $|S_{21}|$. Parameters: d = 0.508mm, ε_r = 2.2, $w_1/d \approx w_2/d \approx w_3/d = 0.717$, $s_1/d \approx s_2/d \approx s_3/d = 0.298$.